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For: CHANNEL ESTIMATION IN A MULTI CARRIER
      TRANSMIT DIVERSITY SYSTEM

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Respectfully submitted,

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If no title is shown please refer to the description.
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Channel estimation in a multi carrier transmit diversity system

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Channel Estimation in a
Multi Carrier Transmit Diversity System

5

BACKGROUND OF THE INVENTION

Technical Field

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The present invention relates to the field of transmit antenna diversity and in particular to a method of estimating channel coefficients in a multi carrier transmit diversity system. The invention also relates to an estimating stage for performing

15   channel estimation operations and to a transceiver of a wireless communications system comprising such an estimating stage.

Discussion of the Prior Art

20   Peak transmission rates in wireless communication systems have steadily increased during the last years. However, peak transmission rates are still limited for example due to path loss, limited spectrum availability and fading.

25   Transmitter diversity is a highly effective technique for combating fading in wireless communications systems. Several different transmit diversity schemes have been proposed. In Li, Y.; Chuang, J.C.; Sollenberger, N.R.: Transmitter diversity for OFDM systems and its impact on high-rate data wireless net-
30   works, IEEE Journal on Selec. Areas, Vol. 17, No. 7, July 1999 the transmit diversity schemes of delay, permutation and space-time coding are exemplarily described. According to the delay approach, a signal is transmitted from a first transmitter antenna and signals transmitted from further transmitter anten-
35   nas are delayed versions of the signal transmitted from the first transmitter antenna. In the permutation scheme, the modulated signal is transmitted from a first transmitter antenna and permutations of the modulated signal are transmitted

from further transmitter antennas. Thus, the signal transmitted
from the transmitter antennas can be derived from a matrix
composed of data words in the form of the modulated signal and
of permutations of the modulated signal. By means of space-time

5   coding a signal is encoded into several data words and each
data word is transmitted from a different transmitter antenna.
During transmission the data words are spread (i.e., multi-
plexed) in the time domain by successively transmitting the
data symbols of each data word over a single carrier frequency.

10

Another transmit diversity scheme is described in US 6,088,408.
According to this transmit diversity scheme data are coded in
the form of matrices and transmitted as individual data blocks.
Each data block comprises several data words and each data word

15   contains data symbols derived from an input data signal. During
transmission of the data blocks, the individual data words are
spread in the time domain. Therefore, the transmit diversity
scheme described in US 6,088,408 can be referred to as space-
time block coding (STBC). The main features of STBC are that

20   each data symbol is transmitted from each transmit antenna and
that the antenna signals of different transmit antennas are
orthogonal to each other. Orthogonal STBC data blocks can be
designed for an arbitrary number of transmit antennas.

25   A further transmit diversity scheme for a multi carrier system
is space-frequency block coding (SFBC). By means of space-
frequency block coding a signal is encoded into individual data
blocks comprising several data words and each data word is
spread (i.e., multiplexed) in the frequency domain by transmit-

30   ting the data symbols of each data word on orthogonal frequen-
cies, i.e. orthogonal subcarriers. An exemplary scheme for
space-frequency block coding is described in Lee K. F. et al:
"A space-frequency transmitter diversity technique for OFDM
systems", Globecom '00 - IEEE. Global Telecommunications Con-

35   ference. Conference record (Cat. No. 00 CH37137), Proceedings
of global Telecommunications conference, San Francisco, CA,
USA, 27 Nov. - 1 Dec. 2000, pages 1473 - 1477 vol. 3.

An important feature on a receiving side of a multi carrier transmit diversity system is a characterization of the individual transmit channels which enables an efficient demodulation of the receive signal. Consequently, a channel estimation has to be performed as described in Li, Y.; Chuang, J.C.; Sollenberger, N.R.: Transmitter diversity for OFDM systems and its impact on high-rate data wireless networks, IEEE Journal on Selec. Areas, Vol. 17, No. 7, July 1999 and in US 6,088,408.

Departing from the various channel estimation approaches known in the art there is a need for a more accurate method of estimating channel coefficients in a multi carrier system operating in accordance with a code matrix based transmit diversity scheme in which a data content of the code matrix is multiplexed in the frequency domain. There is also a need for an estimating stage for performing the corresponding estimating method and for a transceiver comprising such an estimating stage .

BRIEF DESCRIPTION OF THE INVENTION

The existing need is satisfied by a method of estimating channel coefficients in a multi carrier system operating in accordance with a block-code based transmit diversity scheme in which a data content, e.g. data words, of a code matrix is multiplexed in a frequency domain, comprising determining a phase ramp in the frequency domain or an equivalent thereof in the time domain, the phase ramp or the equivalent thereof being comprised within a receive signal after timing synchronization, processing the receive signal to remove the phase ramp or the equivalent thereof, and estimating the channel coefficients on the basis of the processed receive signal. The phase ramp may be introduced by the transmit channel, by a receive or transmit filter, or during timing synchronization.

The channel estimation method of the invention is not restricted to a specific block-coding scheme as long as the utilized transmit diversity scheme enables to generate from a

data signal code matrices which can be multiplexed in the
frequency domain. The code matrices preferably have the form of
data blocks comprising data words, wherein each data word
contains data symbols derived from the data signal. For exam-

5   ple, the transmit diversity schemes of SFBC and of permutation
in the frequency domain allow to generate such code matrices.

According to the invention it is not necessary that the em-
ployed transmit diversity scheme uses pure frequency domain

10   multiplexing. For example, a transmit diversity scheme which
uses STBC and SFBC in an alternating manner, i.e. which alter-
natingly multiplexes in the time domain and in the frequency
domain, may also be used. In such a case the inventive method
of estimating channel coefficients may be activated each time

15   the system switches from STBC to SFBC and de-activated each
time the system switches from SFBC to STBC.

The channel estimation method according to the invention does
not require that the transmit diversity scheme guarantees full

20   transmit diversity and orthogonality. In other words, the
invention does not necessitate that each data symbol comprised
within a data signal is transmitted on a different frequency.
Nonetheless, a preferred embodiment of the invention comprises
the feature of full transmit diversity and orthogonality.

25

Moreover, the invention is not restricted to any number of
transmit and receive antennas. Preferably, the code matrix is
chosen such that the number of data words per data block equals
the number of transmit antennas. If more than one receive

30   antenna is provided, the receive diversity scheme of maximum-
ratio combining can be applied. However, other receive diver-
sity schemes may be used as well.

Various alternatives for determining the phase ramp in the

35   frequency domain or the equivalent thereof in the time domain
exist. For example, the phase ramp or the equivalent thereof
can be determined by way of estimation. Estimates for the phase
ramp or the equivalent thereof are preferably derived by linear

regression. Instead of or in addition to an estimation of the
phase ramp or of the equivalent thereof, calculation or meas-
urement approaches may be used.

5   Timing synchronization, which might be one of several reasons
for the phase ramp comprised within the receive signal, can be
performed in many ways. Possibly, timing synchronization is
performed such that intersymbol interference is minimized.
Thus, the timing instant which is optimum with respect to
10   minimum interference power can be chosen for system timing
purposes.

Determining the phase ramp or the equivalent thereof and remov-
ing the determined phase ramp or the equivalent thereof from
15   the receive signal can be performed either in the frequency
domain or in the time domain. It is also possible to perform
one of these two steps in the frequency domain and the other
step in the time domain. In the time domain, the equivalent of
the phase ramp will be a delay. This delay can be determined
20   and removed from the receive signal prior to an estimation of
the channel coefficients.

Removal of the phase ramp or the equivalent thereof may be
performed at various locations and at different points in time.
25   For example, when after timing synchronization the receive
signal is split and fed into a channel estimation branch on the
one hand and a demodulation branch on the other hand, removal
of the phase ramp or the equivalent thereof may be effected
either in the channel estimation branch or prior to splitting
30   of the receive signal.

It has been pointed out above that the channel coefficients are
estimated using a processed receive signal including no or at
least a reduced phase ramp. However, once the channel coeffi-
35   cients have been estimated, the phase ramp or the equivalent
thereof, which has previously been determined, may be intro-
duced into the estimated channel coefficients again. Such an
introduction of the phase ramp or the equivalent thereof into

the estimated channel coefficients is advantageous if the
receive signal, which is to be demodulated using the estimated
channel coefficients, also comprises the phase ramp or the
equivalent thereof. This might be the case for example when the
5   phase ramp removal is performed in the channel estimation
branch.

The above method can be implemented both as a computer program
product comprising program code portions for performing the
10 method and as a hardware solution. The hardware solution is
constituted by an appropriately configured estimating stage for
estimating channel coefficients in a multi carrier transmit
diversity system operating in accordance with a block-coding
scheme. The estimating stage has a unit for determining a phase
15 ramp in the frequency domain or an equivalent thereof in the
time domain comprised within a receive signal after timing
synchronization, a unit for processing the receive signal to
remove the phase ramp or the equivalent thereof, and a unit for
estimating the channel coefficients using the processed receive
20 signal.

In a signal path behind a timing synchronization unit a node
might be arranged for splitting a common signal path into a
channel estimation branch and a demodulation branch. The proc-
25 essing unit may be arranged either in the channel estimation
branch or in the common signal path prior to the node. Prefera-
bly, the estimating stage further comprises a unit for intro-
ducing the phase ramp or the equivalent thereof into the
estimated channel coefficients.
30

Many transmit diversity schemes require constant or at least
approximately constant channel parameters, i.e. channel phase
and channel amplitude in the frequency domain, during transmis-
sion of one data word. Since the data words are to be multi-
35 plexed in the frequency domain, a comparatively large coherence
band width is required. This means that the relation

$$B_C \gg N/T \qquad\qquad (1)$$

has to be fulfilled at least approximately, wherein $B_c \approx 1/\tau_{rms}$
is the channel cherence bandwidth, N is the number of data
symbols per data word, T is the duration of one of the data
symbols, i.e. the duration of one time slot, and τ_{rms} is the
root mean square of the delay spread of the channel impulse
response. A comparatively large coherence bandwidth requires
that the channel coefficients of N adjacent subcarriers have to
be almost constant.

The data signal from which the one or more code matrices, i.e.
data blocks, are generated can have any format. According to a
preferred embodiment, the data signal has the format of a
sequence of discrete information symbols. For example, the data
signal may have the structure of vectors, each vector compris-
ing a predefined number of information symbols. The nature of
the information symbols may depend on the specific wireless
communication system in which the multiplexing method according
to the invention is used. Many wireless communication systems
employ different types of information symbols for different
purposes. For example, some wireless communication systems use
data signals which comprise a preamble, one or more user data
sections or both a preamble and one or more user data sections.
Usually, the preamble has a predefined structure and enhances
functions like channel estimation, frequency synchronization
and timing synchronization.

The code matrix may be derived from the data signal in various
ways dependent on the transmit diversity scheme which is actu-
ally used. If, for example, the transmit diversity scheme of
permutation is used, the data symbols contained in the data
words of the code matrix are permutations of information sym-
bols comprised within the original data signal. As a further
example, if the transmit diversity scheme of SFBC is used, the
data symbols contained in the data words of the code matrix are
obtained from the information symbols comprised within the
original data signal by means of permutation and basic arithme-
tic operations, such as negation and complex conjugation.

BRIEF DESCRIPTION OF THE DRAWINGS

Further advantages of the invention will become apparent by
reference to the following description of preferred embodiments
of the invention in the light of the accompanying drawings, in
which:

Fig. 1   shows a data signal in the form of a physical
              burst to be processed in accordance with the in-
              vention;

Fig. 2   shows the structure of an OFDM symbol comprising
              a cyclic prefix;

Fig. 3   is a block diagram of a transmitter stage of a
              transceiver for wireless communication;

Fig. 4   shows several modulation schemes defined in the
              HIPERLAN/2 standard;

Fig. 5   shows a block code encoder of the transceiver
              depicted in Fig. 3;

Fig. 6   shows a configuration of a transmit antenna
              diversity system;

Fig. 7   is a schematic diagram of multiplexing data
              words in the time domain;

Fig. 8   is a schematic diagram of multiplexing data
              words in the frequency domain.

Fig. 9   is a schematic diagram of several components of
              a receiver stage of a transceiver for wireless
              communication;

Fig. 10A, 10B   show the introduction of a delay into a receive
                signal during timing synchronization;

Fig. 11   is a schematic diagram of a receiver stage
                comprising a first embodiment of an estimating
                stage according to the invention; and

Fig. 12   is a schematic diagram of a receiver stage
                comprising a second embodiment of an estimating
                stage according to the invention.

DESCRIPTION OF PREFERRED EMBODIMENTS

Although the present invention can be used in any multi carrier
transmit diversity system which employs a transmit diversity
scheme allowing to generate data blocks having a structure
similar to e.g. a SFBC code matrix and to perform multiplexing
in the frequency domain, the following description of preferred
embodiments is exemplarily set forth with respect to a multi
carrier system which employs orthogonal frequency division
multiplexing (OFDM) and which alternately utilizes STBC and
SFBC for generating data blocks from a data signal and for
multiplexing the generated data blocks.

The exemplary multi carrier system is derived from the European
wireless local area network (WLAN) standard HIgh PErformance
Radio Local Area Network type 2 (HIPERLAN/2). HIPERLAN/2 sys-
tems are intended to be operated in the 5 GHz frequency band.
Up to now, the HIPERLAN/2 system and many other wireless commu-
nications systems do not support transmit diversity in spite of
the fact that transmit diversity would improve the transmission
performance and reduce negative effects of fast fading like
Rayleigh fading. A system overview of HIPERLAN/2 is given in
ETSI TR 101 683, Broadband Radio Access Networks (BRAN);
HIPERLAN Type 2; System Overview, V1.1.1 (2000-02) and the
physical layer of HIPERLAN/2 is described in ETSI TS 101 475;
Broadband Radio Access networks (BRAN); HIPERLAN Type 2; Physi-
cal (PHY) Layer, V1.1.1 (2000-04). The multi carrier scheme of

OFDM, which is specified in the HIPERLAN/2 standard, is very
robust in frequency selective environments.

In Fig. 1 a typical physical burst of HIPERLAN/2 is illus-
trated. The physical burst comprises a preamble consisting of
preamble symbols and a user data section consisting of user
data symbols. In HIPERLAN/2 five different physical bursts are
specified. Three of the physical bursts have a different pream-
ble each and the two remaining bursts have a further preamble
in common. The last three preamble symbols constitute a peri-
odic structure which is identical for all preamble types. This
periodic structure consists of a short OFDM symbol C32 of 32
samples followed by two identical regular OFDM symbols C64 of
64 samples. The short OFDM symbol C32 is a cyclic prefix which
is a repetition of the second half of one of the C64 OFDM
symbols. The so-called C-preamble depicted in Fig. 1 is used in
HIPERLAN/2 for channel estimation, frequency synchronization
and timing synchronization. The periodic structure within the
C-preamble is necessary in order to allow the use of synchroni-
zation algorithms with comparatively low complexity.

The user data section of the physical burst depicted in Fig. 1
comprises a variable number N_{SYM} of OFDM symbols required to
transmit a specific protocol data unit (PDU) train. Each OFDM
symbol of the user data section consists of a cyclic prefix and
a useful data part. The cyclic prefix consists of a cyclic
continuation of the useful data part and is inserted before it.
Thus, the cyclic prefix is a copy of the last samples of the
useful data part as depicted in Fig. 2.

The length of the useful data part of the physical burst shown
in Fig. 1 is equal to 64 samples and has a duration of 3,2 μs.
The cyclic prefix has a length of either 16 (mandatory) or 8
(optional) samples and a duration of 0,8 μs or 0,4 μs, respec-
tively. Altogether, a OFDM symbols has a length of either 80 or
72 samples corresponding to a symbol duration of 4,0 μs or 3,6
μs, respectively. An OFDM symbol therefore has an extension in
the time domain. A OFDM symbol further has an extension in the

frequency domain. According to HIPERLAN/2, a OFDM symbol ex-
tends over 52 subcarriers. 48 subcarriers are reserved for
complex valued subcarrier modulation symbols and 4 subcarriers
are reserved for pilots.

5

In Fig. 3, the physical layer of a transmitter stage 10 of a
transceiver for wireless communication is illustrated. The
transmitter stage 10 comprises a scrambler 12, an FEC coding
unit 14, an interleaving unit 16, a mapping unit 18, an OFDM

10   unit 20, a burst forming unit 22, a block code encoder 24, a
multiplexer 26, a radio transmitter 30 and a control unit 32.
The block code encoder 24 and the multiplexer 26 together form
an encoder/multiplexer unit 28.

15   The transmitter stage 10 depicted in Fig. 1 receives as input
signal a PDU train from a data link control (DLC). Each PDU
train consists of information bits which are to be framed into
a physical burst, i.e. a sequence of OFDM symbols to be en-
coded, multiplexed and transmitted.

20

Upon receipt of a PDU train the transmission bit rate within
the transceiver is configured by choosing an appropriate physi-
cal mode based on a link adaption mechanism. A physical mode is
characterized by a specific modulation scheme and a specific

25   code rate. In the HIPERLAN/2 standard several different coher-
ent modulation schemes like BPSK, QPSK, 16-QAM and optional 64-
QAM are specified. Also, for forward error control, convolu-
tional codes with code rates of 1/2, 9/16 and 3/4 are specified
which are obtained by puncturing of a convolutional mother code

30   of rate 1/2. The possible resulting physical modes are depicted
in Fig. 4. The data rate ranging from 6 to 54 Mbit/s can be
varied by using various signal alphabets for modulating the
OFDM subcarriers and by applying different puncturing patterns
to a mother convolutional code.

35

Once an appropriate physical mode has been chosen, the N_{BPDU}
information bits contained within the PDU train are scrambled
with the length-127 scrambler 12. The scrambled bits are then

output to the FEC coding unit 14 which encodes the N_{BPDU} scrambled PDU bits according to the previously set forward error correction.

5   The encoded bits output by the FEC coding unit 14 are input into the interleaving unit 16 which interleaves the encoded bits by using the appropriate interleaving scheme for the selected physical mode. The interleaved bits are input into the mapping unit 18 where sub-carrier modulation is performed by
10   mapping the interleaved bits into modulation constellation points in accordance with the chosen physical mode. As mentioned above, the OFDM subcarriers are modulated by using BPSK, QPSK, 16-QAM or 64-QAM modulation depending on the physical mode selected for data transmission.

15

The mapping unit 18 outputs a stream of complex valued subcarrier modulation symbols which are divided in the OFDM unit in groups of 48 complex numbers. In the OFDM unit a complex base band signal is produced by OFDM modulation as described in ETSI
20   TS 101 475, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04).

The complex base band OFDM symbols generated within the OFDM unit 20, where pilot subcarriers are inserted, are input into
25   the physical burst unit 22, where an appropriate preamble is appended to the PDU train and the physical burst is built. The physical burst produced by the physical burst unit 22 has a format as depicted in Fig. 1. The physical burst unit 22 thus outputs a sequence of complex base band OFDM symbols in the
30   form of the physical burst to the block code encoder 24.

The function of the block code encoder 24 is now generally described with reference to Fig. 5. In general, the block code encoder 24 receives an input signal in the form of a sequence of vectors $\mathbf{X} = [X_1 X_2 ... X_K]^T$ of the length K. The block code
35   encoder 24 encodes each vector $\mathbf{X}$ and outputs for each vector $\mathbf{X}$ a data block comprising a plurality of signal vectors $\mathbf{C}^{(1)}, \mathbf{C}^{(2)} ..., \mathbf{C}^{(M)}$ as depicted in Fig. 5. Each signal vector $\mathbf{C}^{(1)}$,

$C^{(2)}..., C^{(M)}$ corresponds to a single data word. Thus, the data block generated from the vector X comprises M data words, wherein M is the number of transmitter antennas.

5 Each data word $C^{(i)}$ with i = 1...M comprises N data symbols, i.e. each data word $C^{(i)}$ has a length of N. The value of N cannot be freely chosen since a code matrix C spanned by the data words $C^{(i)}$ has to be orthogonal in this embodiment. Several examples for data blocks in the form of orthogonal code matri-
10 ces C are described in US 6,088,408, herewith incorporated by reference. In the block coding approach described in the present embodiment, all data symbols c_j^i of the code matrix C are derived from the components of the input vector X and are simple linear functions thereof or of its complex conjugate.

15

If a receive signal vector Y at one receive antenna is denoted by $Y = [Y_1 Y_2 ... Y_N]^T$, the relationship between Y and the code matrix C is as follows:

20

$$\begin{bmatrix} Y_1 \\ Y_2 \\ ... \\ Y_N \end{bmatrix} = \begin{bmatrix} c_1^{(1)} & c_1^{(2)} & ... & c_1^{(M)} \\ c_2^{(1)} & ... & & c_2^{(M)} \\ ... & & ... & ... \\ c_N^{(1)} & c_N^{(2)} & ... & c_N^{(M)} \end{bmatrix} \cdot \begin{bmatrix} h^{(1)} \\ h^{(2)} \\ ... \\ h^{(M)} \end{bmatrix} \qquad (2)$$

where $h^{(i)}$ represents the channel coefficient of the channel from the i-th transmit antenna to the receive antenna. A generalization to more receive antennas is straightforward.

25 In the following, examples of possible block code matrices for two and three transmitter antennas, respectively, are discussed in more detail. The configuration of a wireless communications system with two transmit antennas and one receive antenna is depicted in Fig. 6. The wireless communications system of Fig.
30 6 comprises two transmit channels, each transmit channel being characterized by a specific channel coefficient $h^{(i)}$ with i = 1,2.

In the case of two transmit antennas depicted in Fig. 6 one
possible block code matrix **C** with a code rate R = 1 is :

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_1^* \end{bmatrix} \tag{3}$$

For three transmit antennas one possible block code matrix **C**
5   with a code rate R = 0,5 is:

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 & X_3 \\ -X_2 & X_1 & -X_4 \\ -X_3 & X_4 & X_1 \\ -X_4 & -X_3 & X_2 \\ X_1^* & X_2^* & X_3^* \\ -X_2^* & X_1^* & -X_4^* \\ -X_3^* & X_4^* & X_1^* \\ -X_4^* & -X_3^* & X_2^* \end{bmatrix} \tag{4}$$

The code rate R is defined as the ratio of the length K of the
10   input vector **X** and the length N of each code word **C**$^{(i)}$:

$$R = K/N \tag{5}$$

As can be seen from Fig. 5, the block code encoder 24 outputs
15   for each data signal in the form of a vector **X** a data block in
the form of a code matrix **C**. The data block output by the block
code encoder 24 is input into the multiplexer 26 which multi-
plexes the data words (vectors **C**$^{(i)}$) of each data block in
accordance with an externally provided control signal in the
20   frequency domain. The control signal is generated by the con-
trol unit 32.

In the multi carrier scheme OFDM, the output of the block code
encoder 24 is modulated onto subcarriers which are orthogonal
25   to each other. There exist essentially two possibilities to
multiplex a data block comprising individual data words in an
OFDM system. According to a first possibility depicted in Fig.

7, the data words of a specific data block could be extended in
the time direction (STBC), i.e. multiplexed in the time domain.
According to a second possibility, which his employed for the
present invention, the data words of a data block are extended
5   in the frequency direction as depicted in Fig. 8 (SFBC).

As can be seen from Figs. 7 and 8, the individual data words of
a data block are transmitted from different transmit antennas.
According to the multiplexing scheme of Fig. 8, an individual
10 data block is spread over N subcarriers and is transmitted
during a time interval of T.

The encoded and multiplexed output signal of the encoder/multi-
plexer unit 28 is input into the radio transmitter 30. The
15 radio transmitter 30 performs radio transmission over a plural-
ity of transmit antennas by modulating a radio frequency car-
rier with the output signal of the encoder/multiplexer unit 28.

The transceiver with the transmitter stage 10 of Fig. 3 further
20 comprises a receiver stage not depicted in Fig. 3. The receiver
stage has a physical layer with components for performing the
inverse operations of the components depicted in Fig. 3. For
example, the receiver stage comprises a descrambler, a FEC
decoding unit, a demultiplexer/decoder unit with a demulti-
25 plexer and a block code decoder, etc. Fig. 9 shows some compo-
nents of such a receiver stage 40.

As becomes apparent from Fig. 9, a receive signal vector **Y**
received via a receive antenna not depicted in Fig. 9 is fed
30 into a timing synchronization unit 42, which performs timing
synchronization with the object to find an optimum timing
instant minimizing intersymbol interference among the received
data symbols. The output signal of the timing synchronization
unit 42 is concurrently fed into a channel estimating unit 44
35 and a demodulator 46. The channel estimating unit 44 estimates
the channel coefficients $h^{(i)}$ on the basis of the output signal
of the timing synchronization unit 42. The channel coefficients
estimated by the channel estimating unit 44 are then passed on

to the demodulator 46 which performs demodulation of the receive signal using the estimated channel coefficients.

In the following, the operation of the channel estimating unit 44 is exemplarily described for the case that two transmit antennas and one receive antenna are utilized (Fig. 6). In this case encoding of the data signal can be performed on the basis of the above block code matrix (3) and the receive signal vector can be written as $\mathbf{Y} = [Y_j \ Y_{j+1}]^T$. The index j denotes a specific frequency f_j.

On frequency f_j, X_i is transmitted from the first transmit antenna and X_{i+1} is transmitted from the second transmit antenna. On an adjacent frequency f_{j+1}, $-X^*_{i+1}$ is transmitted from the first transmit antenna and X^*_i is transmitted from the second transmit antenna. The individual components Y_j and Y_{j+1} of the receive signal vector $\mathbf{Y}$ can thus be written as

$$Y_j = X_i \cdot h^{(1)}(z_j) + X_{i+1} \cdot h^{(2)}(z_j) + n_j$$

$$Y_{j+1} = -X^*_{i+1} \cdot h^{(1)}(z_{j+1}) + X^*_i \cdot h^{(2)}(z_{j+1}) + n_{j+1}$$

$$(6)$$

The variable z_j denotes the transmit frequency f_j. Therefore, $h^{(i)}(z_j)$ is the coefficient of the channel between the transmit antenna i = 1,2 and the receive antenna for a data symbol transmitted via frequency f_j (SFBC). The term n_j denotes the white gaussian noise at frequency f_j.

For the case that SFBC is applied ($z_j = f_j$) and the coherence bandwidth B_c is relatively large, i.e. if relation (1) is fulfilled, the following assumptions are valid

$$h^{(1)}(z_j) = h^{(1)}(z_{j+1}) = h^{(1)}$$

$$h^{(2)}(z_j) = h^{(2)}(z_{j+1}) = h^{(2)}$$

$$(7)$$

This means that if the coherence bandwidth B_c is relatively large equations (6) become

$$Y_j = X_i \cdot h^{(1)} + X_{i+1} \cdot h^{(2)} + n_j$$

5

$$Y_{j+1} = -X^*_{i+1} \cdot h^{(1)} + X^*_i \cdot h^{(2)} + n_{j+1}$$

$$(8)$$

Equations (8) can be written in terms of the receive signal
10   vector $\mathbf{Y}$ and a data matrix $\mathbf{Z}$, which is equivalent to the code matrix $\mathbf{C}$, as

$$\mathbf{Y} = \begin{bmatrix} Y_j \\ Y_{j+1} \end{bmatrix} = \begin{bmatrix} X_i & X_{i+1} \\ -X^*_{i+1} & X^*_i \end{bmatrix} \cdot \begin{bmatrix} h^{(1)} \\ h^{(2)} \end{bmatrix} + \begin{bmatrix} n_j \\ n_{j+1} \end{bmatrix} = \mathbf{Z} \cdot \mathbf{H} + \mathbf{N}$$

$$(9)$$

15

In order to provide an estimate for the channel coefficients
$h^{(1)}$ and $h^{(2)}$, the receive signal vector $\mathbf{Y}$ is multiplied with the
Hermitian $\mathbf{Z}^H$ of the known data matrix $\mathbf{Z}$. The content of the
data matrix $\mathbf{Z}$ corresponds to a standardized preamble portion
20   which is known to the transceiver. The multiplication of $\mathbf{Z}^H$ and
$\mathbf{Y}$ yields

$$\mathbf{Z}^H \cdot \mathbf{Y} = \mathbf{Z}^H \cdot \mathbf{Z} \cdot \mathbf{H} + \mathbf{Z}^H \cdot \mathbf{N} = \hat{\mathbf{H}}$$

$$(10)$$

25

Since $\mathbf{Z}$ is a scaled unitary matrix, i.e.

$$\mathbf{Z}^{-1} = \frac{1}{\det(Z)} \mathbf{Z}^H,$$

$$(11)$$

30

the channel coefficients in equation (10) are separated. The
estimated channel coefficients comprised within $\hat{\mathbf{H}}$ are passed
from the channel estimating circuit 44 to the demodulator 46.

It has been found that the estimated channel coefficients thus
obtained deviate from the actual channel coefficients. One
reason therefor is the fact that after timing synchronization a
phase ramp in the frequency domain is present in the receive
5   signal Y to be processed by the channel estimating unit 44 and
the demodulator 46. The origins and the influence of the phase
ramp present in the receive signal Y after timing synchroniza-
tion will now be described in more detail.

10   It has been mentioned above that the timing synchronization
unit 42 performs timing synchronization with the object of
minimizing intersymbol interference. In many cases there are
multiple optimum timing instants where no intersymbol interfer-
ence occurs. For example, this applies if the cyclic prefix of
15   an OFDM symbol is longer than the channel impulse response. In
this situation of multiple timing instants, which are all
optimum with respect to minimizing intersymbol interference,
the actual timing position is chosen dependent only on the
currently involved noise samples.
20

In Figs. 10A and 10B the phase of the channel transfer function
H(f) of a one tap channel for different timing positions mini-
mizing the intersymbol interference is shown. The position "0"
in the time domain refers to the timing synchronization at the
25   receiver stage 40 relative to a FFT window of the receiver
stage 40.

As becomes apparent from Fig. 10A, there is no phase ramp
comprised in the phase arc(H(f)) of the channel transfer func-
30   tion H(f) when the one tap is at timing position "0". Regarding
OFDM, however, it is very likely that the obtained timing
synchronization introduces a delay Δt relative to the FFT
window of the receiver stage 40. This situation is depicted in
Fig. 10B. The reason for this delay Δt is not the processing
35   time of the timing synchronization unit 42, but the fact that
the optimum timing instant is chosen such that the intersymbol
interference is minimized. The delay Δt can therefore also be
referred to as a systematic delay. The consequence of the delay

Δt introduced during timing synchronization is a systematic phase ramp, i.e. a non-flat phase of channel transfer function $H(f)$. This phase ramp is depicted in Fig. 10B on the right.

5   The phase ramp comprised within the receive signal may not only be caused by timing synchronization, but also by the transmit channel or by receive and transmit filters. If the phase ramp is already present in the receive signal prior to timing syn-chronization, the phase ramp will usually decrease as a result
10   of timing synchronization. This applies even if timing synchro-nization is performed with the object of minimizing intersymbol interference. Nevertheless, the probability that there will be no phase ramp in the receive signal after timing synchroniza-tion is comparatively low.

15
Since the channel coefficients are estimated based on a receive signal comprising a phase ramp, the phase ramp will also be present in the estimated channel coefficients. The phase ramp contained in the channel coefficients leads to an increasing
20   interference during the channel estimation when SFBC is ap-plied, i.e. if the data words are multiplexed in the frequency domain. The reason for this increasing interference is the fact that equation (7) will generally not be fulfilled even if relation (1) holds, i.e. even if the coherence bandwidth B_c is
25   relatively large.

Since one reason for the phase ramp is the fact that timing synchronization is performed based on other criteria than minimizing the phase ramp, one could think about implementing
30   timing synchronization with the object of minimizing the phase ramp instead of minimizing the intersymbol interference. In this case the channel estimation will be improved. However, a drawback of this solution would be the fact that the overall performance of the receiver stage 40 will decrease since inter-
35   symbol interference will no longer be minimized.

In order to both minimize intersymbol interference and improve channel estimation, the receiver stage 40 depicted in Fig. 11

is proposed. The improved receiver stage 40 comprises an esti-
mating stage 60 according to a first embodiment of the inven-
tion with a phase ramp estimation unit 48, a phase ramp removal
unit 50, a channel estimating unit 44, and a phase ramp intro-
duction unit 52. In a signal path behind the timing synchroni-
zation unit 42 a node 54 is arranged for splitting a common
signal path 55 into a channel estimating branch 56 and a de-
modulation branch 58. As can be seen from Fig. 11, the channel
estimating unit 44 is arranged in the channel estimating branch
56.

The function of the estimating stage 60 depicted in Fig. 11
will now be described exemplarily for the frequency domain. The
individual steps performed by the estimating stage 60 could
also be performed in the time domain.

As has been explained above, the transmit channel, a receive or
a transmit filter, or the timing synchronization unit 42 may
introduce a phase ramp $\varphi(\omega)$ in the frequency domain or a
delay Δt in the time domain into a receive signal Y. The re-
ceive signal $Y_{\Delta t}$ including this phase ramp $\varphi(\omega)$ is input from
the timing synchronization unit 42 into the phase ramp estima-
tion unit 48 of the estimating stage 60. The phase ramp esti-
mating unit 48 determines by way of linear regression the phase
ramp $\varphi(\omega)$ introduced into the receive signal.

The estimated phase ramp $\varphi_{est}(\omega)$ is input into the phase ramp
removal unit 50 which his arranged in the channel estimation
branch 56. The phase ramp removal unit 50 also receives the
receive signal $Y_{\Delta t}$, which comprises the phase ramp $\varphi(\omega)$, from
the timing synchronization unit 42 and processes this receive
signal $Y_{\Delta t}$ to remove the phase ramp $\varphi(\omega)$ comprised therein.
This is done by multiplying the receive signal $Y_{\Delta t}$ with a fac-
tor $e^{-j\varphi_{est}(\omega)}$.

The phase ramp removal could also be performed in the time
domain, i.e. the delay Δt corresponding to the phase ramp φ

could be compensated. In this case the delay compensation is
equal to a cyclic shift of the samples of each OFDM symbol to
which the correction should be applied.

5   The receive signal $Y_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$ processed by the phase ramp
removal unit 50 is input into the channel estimating circuit
44. The channel estimating unit 44 estimates the channel coef-
ficients $h^{(i)}$ as previously explained with respect to equations
(6) to (11).

10

The estimated channel coefficients $\hat{h}^{(i)}$ output by the channel
estimating unit 44 are input into the phase ramp introduction
unit 52 which further receives the phase ramp $\varphi_{est}(\omega)$ from the
phase ramp estimating unit 48. The phase ramp introduction unit

15   52 introduces the removed estimated phase ramp $\varphi_{est}(\omega)$ again by
multiplying the estimated channel coefficients $\hat{h}^{(i)}$ with
$e^{j\varphi_{est}(\omega)}$. The channel coefficients $\hat{h} \cdot e^{j\varphi_{est}(\omega)}$ output by the
phase ramp introduction unit 52 are input to the demodulator 46
where demodulation of the receive signal $Y_{\Delta t}$ is performed util-

20   izing the channel coefficients.

In Fig. 12, a receiver stage 40 comprising an estimating stage
60 according to a second embodiment of the invention is shown.
The estimating stage 60 comprises a phase ramp estimation unit

25   48, a phase ramp removal unit 50 and a channel estimating unit
44. The phase ramp removal unit 50 is arranged in a common
signal path 55 behind a timing synchronization unit 52 and
prior to a node 54. The node 54 serves for splitting the common
signal path 55 into a channel estimating branch 56 and a de-

30   modulation branch 58.

The function of the estimating stage 60 depicted in Fig. 12
will now be described exemplarily for the frequency domain. The
individual steps performed by the estimating stage 60 could

35   also be performed in the time domain.

The receive signal $Y_{\Delta t}$ including a phase ramp $\varphi(\omega)$ is input
from the time synchronization unit 42 into the phase ramp

estimation unit 48 of the estimating stage 60. The phase ramp
estimating unit 48 determines by way of linear regression the
phase ramp $\varphi(\omega)$ introduced into the receive signal **Y**. The
estimated phase ramp $\varphi_{est}(\omega)$ is then input into the phase ramp

5   removal unit 50 which is arranged in the common signal path 55.
The phase ramp removal unit 50 also receives the receive signal
Y$_{\Delta t}$, which comprises the phase ramp $\varphi(\omega)$, from the timing
synchronization unit 42 and processes this receive signal **Y**$_{\Delta t}$
to remove the phase ramp $\varphi(\omega)$ comprised therein. This is done

10 in a similar way like in the first embodiment by multiplying
the receive signal **Y**$_{\Delta t}$ with a factor $e^{-j\varphi_{est}(\omega)}$.

The receive signal **Y**$_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$ processed by the phase ramp
removal unit 50 is split at a node 54 and fed both into the

15 channel estimation branch 56 and the demodulation branch 58. In
the channel estimation branch 56 the channel estimating unit 44
determines estimated channel coefficients $\hat{h}^{(i)}$ which are input
into the demodulator 46. The demodulator 46 performes demodula-
tion of the processed receive signal **Y**$_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$ utilizing

20 the estimated channel coefficients $\hat{h}^{(i)}$. Since both channel
estimation and demodulation are performed on the basis of the
processed receive signal **Y**$_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$, the phase ramp intro-
duction unit of the first embodiment depicted in Fig. 11 can be
omitted.

25

The embodiment depicted in Fig. 11, according to which removal
of the phase ramp takes place in the channel estimation branch,
can be employed if a transmit diversity scheme is utilized
which multiplexes a preamble portion of a data signal in the

30 frequency domain (e.g. by means of SFBC) and a user data por-
tion of the data signal in the time domain (e.g. by means of
STBC). If however, both the preamble portion and the user data
portion are multiplexed in the frequency domain, the phase ramp
problem outlined above concerns the user data portion also. In

35 this case the embodiment depicted in Fig. 12 can be employed.

The basic concept underlying the invention can be extended to
transmit diversity systems comprising more than two transmit

antennas. A possible further embodiment of the invention is
based on a transmit diversity system comprising three transmit
antennas and operating in accordance with a block-coding scheme
using the code matrix shown in equation (4).

5

CLAIMS

1. A method of estimating channel coefficients (h) in a multi carrier system operating in accordance with a block-code based transmit diversity scheme, in which a data content ($C^{(i)}$) of a code matrix (**C**) is multiplexed in a frequency domain, comprising:

    a)   determining a phase ramp (φ_{est}) in the frequency domain or an equivalent (Δt) thereof in the time domain, the phase ramp (φ_{est}) or the equivalent (Δt) thereof being comprised within a receive signal ($Y_{\Delta t}$) after timing synchronization;

    b)   processing the receive signal ($Y_{\Delta t}$) to remove the phase ramp (φ_{est}) or the equivalent (Δt) thereof; and

    c)   estimating the channel coefficients (h) on the basis of the processed receive signal ($Y_{\Delta t}$).

2. The method of claim 1, wherein the phase ramp (φ_{est}) or the equivalent (Δt) thereof is determined by way of estimation.

3. The method of claim 2, wherein the estimation is performed by linear regression.

4. The method of one of claims 1 to 3, further comprising the step of performing timing synchronization with the object of minimizing intersymbol interference.

5. The method of one of claims 1 to 4, wherein at least one of steps a) and b) is performed in the frequency domain.

6. The method of one of claims 1 to 4, wherein at least one
    of steps a) and b) is performed in a time domain.

7. The method of one of claims 1 to 6, wherein after timing
    synchronization the receive signal ($Y_{\Delta t}$) is split and fed
    into a channel estimation branch (56) on the one hand and
    a demodulation branch (58) on the other hand, and wherein
    the phase ramp (φ_{est}) or the equivalent (Δt) thereof is
    removed in the channel estimation branch (56).

8. The method of one of claims 1 to 6, wherein after timing
    synchronization the receive signal ($Y_{\Delta t}$) is split and fed
    into a channel estimation branch (56) on the one hand and
    a demodulation branch (58) on the other hand, and wherein
    the phase ramp (φ_{est}) or the equivalent (Δt) thereof is
    removed prior to splitting of the receive signal ($Y_{\Delta t}$).

9. The method of one of claims 1 to 7, further comprising
    introducing the phase ramp (φ_{est}) or the equivalent (Δt)
    thereof into the estimated channel coefficients ($\hat{h}$).

10. The method of one of claims 1 to 9, further comprising
    demodulating the receive signal ($Y_{\Delta t}$) utilizing the esti-
    mated channel coefficients ($\hat{h}$).

11. The method of one of claims 1 to 10, wherein the block-
    code based transmit diversity scheme of space-frequency
    block coding (SFBC) or of permutation in the frequency do-
    main is employed.

12. A computer program product comprising program code por-
    tions for performing the steps of one of claims 1 to 11
    when the product is run on a computer.

13. The computer program product of claim 12 stored on a
    computer readable recording medium.

14.   An estimating stage (60) for estimating channel coeffi-
cients (h) in a multi carrier system operating in accor-
dance with a block-code based transmit diversity scheme in
which a data content ($C^{(i)}$) of a code matrix (**C**) is multi-
plexed in a frequency domain, comprising:

a) a unit (48) for determining a phase ramp (φ_{est}) in the
frequency domain or an equivalent (Δt) thereof in the time
domain, the phase ramp (φ_{est}) or the equivalent (Δt)
thereof being comprised within a receive signal ($Y_{\Delta t}$) af-
ter timing synchronization;

b)   a unit (50) for processing the receive signal ($Y_{\Delta t}$)
to remove the phase ramp (φ_{est}) or the equivalent (Δt)
thereof; and

c)   a unit (44) for estimating the channel coefficients
(h) on the basis of the processed receive signal ($Y_{\Delta t}$).

15.   The estimating stage according to claim 14, further com-
prising a node (54) for splitting a signal path (55) after
timing synchronization into a channel estimation branch
(56) on the one hand and a demodulation branch (58) on the
other hand, and wherein the unit (50) for processing the
receive signal ($Y_{\Delta t}$) is arranged in the channel estimation
branch (56).

16.   The estimating stage according to claim 14, further com-
prising a node (54) for splitting a signal path (55) after
timing synchronization into a channel estimation branch
(56) on the one hand and a demodulation branch (58) on the
other hand, and wherein the unit (50) for processing the
receive signal ($Y_{\Delta t}$) is arranged in the signal path (55)
prior to the node (54).

17.   The estimating stage according to claim 14 or 15, further
comprising a unit (52) for introducing the phase ramp

(φ_{est}) or the equivalent (Δt) thereof into the estimated channel coefficients ($\hat{h}$).

18. A transceiver of a wireless communication system comprising a receiver stage (40) with an estimating stage (60) according to one of claims 14 to 17.

ABSTRACT

5   **Channel Estimation in a**
            Multi Carrier Transmit Diversity System

A method and a stage (60) for estimating channel coefficients
in a multi carrier system operating in accordance with a block-
10   code based transmit diversity scheme, in which a data content
of a code matrix is multiplexed in a frequency domain are
described. The method comprises determining a phase ramp (φ_{est})
in the frequency domain or an equivalent (Δt) thereof in the
time domain, the phase ramp (φ_{est}) or the equivalent (Δt)
15   thereof being comprised within a receive signal (**Y**) after
timing synchronization, processing the receive signal ($\mathbf{Y}_{\Delta t}$) to
remove the phase ramp (φ_{est}) or the equivalent (Δt) thereof and
estimating the channel coefficients on the basis of the proc-
essed receive signal.

20

(Fig. 11)

25   6022

30

Fig. 1

cyclic prefix   original OFDM symbol

OFDM symbol
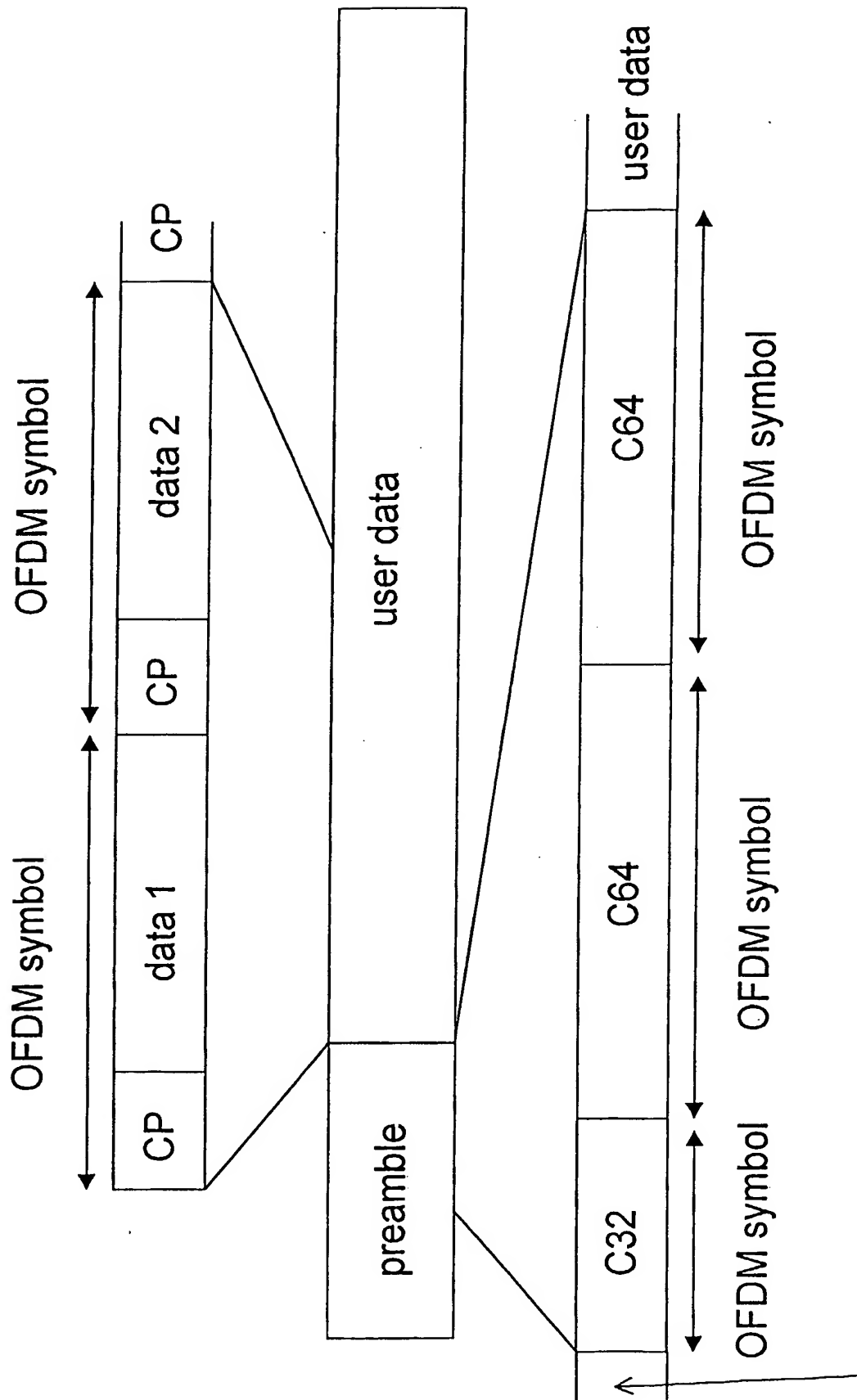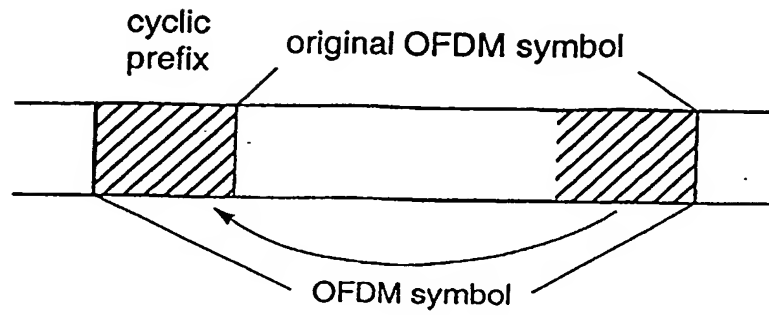
Fig. 2

modulation scheme	code rate	bit rate
BPSK	1/2	6 Mbps
BPSK	3/4	9 Mbps
QPSK	1/2	12 Mbps
QPSK	3/4	18 Mbps
16-QAM	9/16	27 Mbps
16-QAM	3/4	36 Mbps
64-QAM	3/4	54 Mbps

Fig. 4

$$X = [X_1 \, X_2 \, ... \, X_K]^T \longrightarrow \boxed{\begin{array}{c} BC \\ encoder \end{array}}$$

$$\mathbf{C}^{(1)} = [c_1^{(1)} c_2^{(1)} ... c_N^{(1)}]^T$$

$$\mathbf{C}^{(2)} = [c_1^{(2)} c_2^{(2)} ... c_N^{(2)}]^T$$

$$\mathbf{C}^{(M)} = [c_1^{(M)} c_2^{(M)} ... c_N^{(M)}]^T$$

Fig. 5

24

Tx-ant 1 $h^{(1)}$

Tx-ant 2 $h^{(2)}$ Rx-ant

Fig. 6

PDU train from DLC (transmit) → scrambling (12) → FEC coding (14) → interleaving (16) → mapping (18)

OFDM (20) → PHY bursts (22) → Block Code Encoder (24) → Multi-plexer (26)

(28)

Control Unit (32)

radio transmitter (30)

(10)

Fig. 3

1. data symbol (antenna 1) N-th. data symbol (antenna 1)

$$c_1^{(1)}$$

f_{i-1} f_i f_{i+1} f_{i-1} f_i f_{i+1}

1. data symbol (antenna M) N-th. data symbol (antenna M)

$$c_1^{(M)}$$ $$c_N^{(M)}$$

f_{i-1} f_i f_{i+1} f_{i-1} f_i f_{i+1}

Fig. 7

N data symbols (antenna 1)

$$c_1^{(1)} \qquad c_N^{(1)}$$

f_{i-1} f_i f_{i+N}

N data symbols (antenna M)

$$c_1^{(M)} \qquad c_N^{(M)}$$

f_{i-1} f_i f_{i+N}

Fig. 8

Fig. 9

Fig. 10

Fig. 11

phase ramp estimation 48

$\varphi_{est}(\omega)$

Channel estimation 44

$\hat{h}$

56

40

60

42 timing synchronization

γ

$\gamma_{\Delta t}$

55

remove phase ramp 50

$\gamma_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$

54

58

$\gamma_{\Delta t} \cdot e^{-j\varphi_{est}(\omega)}$

46 demodulator

Fig. 12